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Compensation Technique for Capacitive Crosstalk in Continuous-Time Electro-Mechanical Sigma-Delta Modulators

T. Northemann^{a*}, M. Maurer^a, Y. Manoli^{a,b}^a*Fritz Huettinger Chair of Microelectronics, Department of Microsystems Engineering – IMTEK,
University of Freiburg, D-79110 Freiburg, Germany*^b*HSG-IMIT – Institute of Micromachining and Information Technology, D-78052 Villingen-Schwenningen, Germany*

Abstract

This paper presents a readout interface for gyroscopes, where the Coriolis rate signal is converted into a bitstream using a 4th-order continuous-time electro-mechanical sigma-delta modulator. The used gyroscope has separate sense and feedback electrodes allowing full continuous-time readout. A severe drawback arises when a gyroscope with this non-collocated design is employed, since the readout is active during force-feedback. Charge feed-through, caused by parasitic capacitances of the sensor in combination with force-feedback signals, drive the charge integrator front-end into saturation. This paper presents a technique for the compensation of this capacitive crosstalk, enabling stable closed-loop readout interfaces.

© 2011 Published by Elsevier Ltd. Open access under [CC BY-NC-ND license](https://creativecommons.org/licenses/by-nc-nd/4.0/).*Keywords: Sigma-Delta Modulators; Continuous-Time; Crosstalk; Compensation*

1. Introduction

Micromachined angular rate sensors which use the Coriolis effect for detection are used for rollover detection, personal navigation, and image stabilization of cameras. Readout interfaces based on sigma-delta ($\Sigma\Delta$) modulation are widely used [1-5] due to the intrinsic analog-to-digital conversion.

* Corresponding author. Tel.: +49-761-203-7479; fax: +49-761-203-7592.

E-mail address: Thomas.Northemann@imtek.uni-freiburg.de.

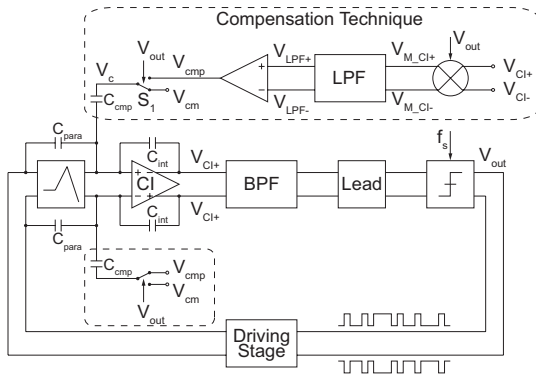


Fig. 1. Electro-mechanical $\Sigma\Delta$ -Modulator with capacitive crosstalk compensation technique.

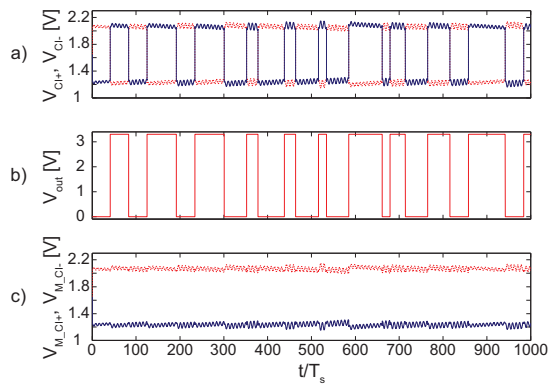


Fig. 2. Signals without compensation technique. The modulator is unstable.

Furthermore, parameter variations of the sensor are suppressed and as the sensor mass is regulated to its zero position, the linearity and dynamic range can be improved as well.

In switched-capacitor (SC) implementations capacitive crosstalk is not an issue since the system is a sampled one. According to [6], a drawback of SC compared to continuous-time (CT) implementations is the use of amplifiers with a higher bandwidth to ensure a fast settling of the switched signals. Furthermore, CT implementations do not suffer from noise folding and have therefore intrinsically a lower noise floor relaxing the amplifier circuit design. In order to use a CT sigma-delta modulator ($\Sigma\Delta$) a compensation technique is necessary to overcome the capacitive crosstalk issue.

2. Sigma-Delta Modulator

Fig. 1 shows the CT electro-mechanical $\Sigma\Delta$ with the proposed compensation technique. For simplicity only the secondary sensor mass is shown. The primary mass is driven into resonance (f_p) with a drive loop as described in [1]. Beside the sensor, a charge integrator (CI), a bandpass filter (BPF), a lead compensator for loop stability, a single-bit quantizer which is sampled with $f_s = 8 \times f_p$, and a 2-level driving stage are used.

For the $\Sigma\Delta$ a feedforward architecture is advantageous, if the sensing element is included in the loop. Compared to distributed feedback architectures, sensor variations do not have an impact on the position of the zeros of the loop filter [4]. As there is no access to the velocity node of the gyroscope a lead compensator is needed to guarantee loop stability. This lead compensator replaces the missing feedforward signal path. The noise-shaping filter is implemented as a bandpass rather than a low-pass filter. Compared to low-pass $\Sigma\Delta$, bandpass solutions can use much lower sampling frequencies relaxing the circuit requirements in terms of power consumption [5], [7]. Beside the filter which is designed according to [8], a single-bit quantizer is used due to its inherent linearity compared to multi-bit solutions.

The 2-level force-feedback signal is applied to the gyroscope. The differential output signal of the charge integrator represents the position of the sensors mass. Without any parasitic capacitances this output signal would be located symmetrically around the common-mode value ($V_{cm} = 1.65V$) as the sensor mass is regulated to its zero position.

Due to the unknown parasitic capacitances C_{para} (in this simulation set to 200fF) a differential charge is generated on the input nodes of the charge integrator driving the output into saturation, Fig. 2(a). As a result the modulator is unstable prohibiting any rate detection.

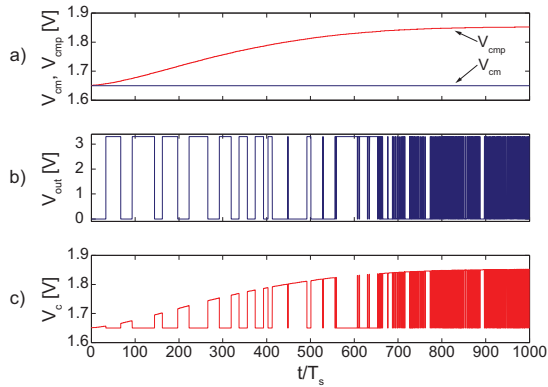


Fig. 3. Signals with compensation technique. Depending on V_{out} , a compensation charge is generated onto the input nodes of the charge integrator using V_c and the capacitor C_{cmp} .

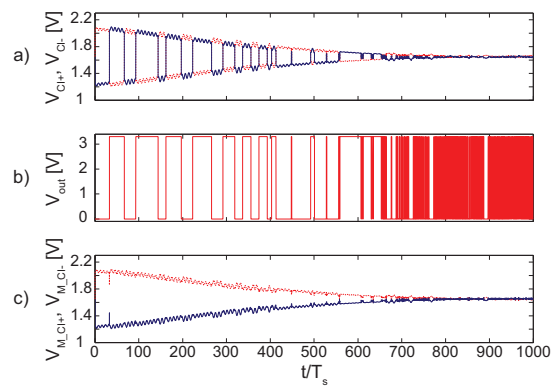


Fig. 4. Signals with the compensation technique. The modulator is stable after $650/T_s$.

3. Compensation Technique

Fig. 1 shows beside the sigma-delta modulator the proposed compensation technique. The additional circuitry consists of a passive mixer as presented in [9], a passive 1st-order low-pass filter, a differential-to-single ended amplifier, two switches, and two compensation capacitors C_{cmp} .

The idea is to apply a compensation charge onto the input nodes of the charge integrator which is equal to the charge caused by the parasitic sensor capacitors but with opposite sign. This is done with the compensation capacitor C_{cmp} and a voltage step V_c which is in anti-phase to the digital output signal V_{out} and therefore to the force feedback signal. This voltage step is defined by the common-mode potential V_{cm} and a compensation voltage V_{cmp} . In order to obtain the exact compensation voltage a control loop is used.

Applying the CI output (Fig. 2a) signals to the mixer, which is controlled by the comparator output V_{out} of the $\Sigma\Delta$ (Fig. 2b), results in Fig. 2(c). In case there is no capacitive crosstalk the difference would be zero in average. Therefore a low-pass filter is necessary before applying these signals to the differential-to-single ended amplifier. Applying a compensation charge onto the CI input nodes with opposite sign reduces this difference again to zero. This is achieved with the compensation capacitor ($C_{cmp}=4\text{pF}$) and the provided compensation signal V_{cmp} , Fig. 3(a). Depending on V_{out} (Fig. 3b) either V_{cm} or V_{cmp} is applied to C_{cmp} via switch S_1 in Fig. 3(c), generating a charge with opposite sign. Depending on the capacitive crosstalk, the compensation signal V_{cmp} can be higher or lower compared to the common-mode potential V_{cm} . A rough estimation of the parasitic sensor capacitors is required since the compensation signal is limited to the power supply. The parasitic charge caused by C_{para} and the force-feedback signal must be within the compensation range. Accordingly C_{cmp} has to be chosen.

Fig. 4 shows the same signals as in Fig. 2. This time with the compensation technique activated. In Fig. 4(a) it can be seen that the saturated signals reduce down to the common-mode level as the compensation signal V_{cmp} increases to its final value, see Fig. 3(a). Or in other word, the difference of the low-pass filtered differential signal of the charge integrator reduces to a residual error defined by the open-loop gain of the differential-to-single ended amplifier. Fig. 4 shows that after 650 samples the modulator is stable.

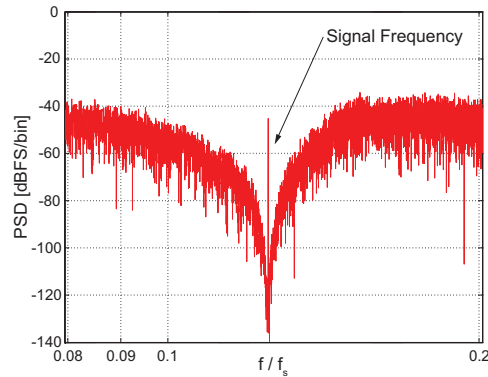


Fig. 5. Simulated spectrum of the 4th-order $\Sigma\Delta$ M sampled with $8 \times f_p$. The applied rate signal is $10^\circ/\text{s}$.

4. Conclusion

With this compensation technique it is possible to implement electro-mechanical $\Sigma\Delta$ modulators in continuous-time techniques. The capacitive crosstalk signal can be compensated with the proposed control loop. The additional hardware effort is small and the additional power consumption of this design is less than $6.6\mu\text{W}$.

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